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20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

This report contains the results of an investigation to quantitatively determine mutual coupling between microstrip radiators and to determine operating parameters of circular ring microstrip antennas which are particularly useful for microstrip array applications.

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AN INVESTIGATION OF THE EFFECTS OF MUTUAL COUPLING ON THE PERFORMANCE OF MICROSTRIP RADIATION

FINAL REPORT

James W. Mink

U.S. ARMY RESEARCH OFFICE

Contract number DAAG2979G0023

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BRIEF OUTLINE OF RESEARCH FINDINGS

As system requirements dictate that antenna arrays achieve extremely low side lobe levels and that beam formation is well known, the effect of mutual coupling between array elements cannot be ignored. This is true even if the coupling between elements is small. At the time this investigation was initiated quantative mutual coupling data was not available (either experimental or theoretical) for use of the microstrip array designer. Hence, one was forced to employ expensive cut-and-try techniques to achieve the final array design.

The goal of this investigation was to quantitatively determine mutual coupling between microstrip radiators and to determine operating parameters of circular ring microstrip antennas which are particularly useful for microstrip array applications.

To achieve the first goal a transmission line model of a pair of coupled rectangular microstrip radiators was developed. This model was then programed on the NCSU IBM 370 computer. The model can handle two basic geometries the H-plane or parallel coupled and te E-plane or gap coupled radiators.

Since microstrip radiators operating at microwave frequencies are normally operated in the dominant TEM mode, and since the spaceing between elements is usually small, a transmission line model is appropriate. Parallel coupling is treated by the well known even/odd mode analysis used in microstrip filter and coupler analysis. Via a four port network formulation, the S-parameters of the coupled pair of radiators is obtained. Gap coupling is modeled by a pi-equivalent network of capacitances. These capacitances are obtained from a variational method of analysis in which the charge distribution on each microstrip radiator is determined.

Results of this investigation has shown that substatial variations in input impedance of the microstrip radiator as the seperation between elements decreases and as the dielectric constant of the substrate material increases. This is in accordance with the increasing non-radiative near-field coupling as shown in appendix 1.

The second investitation was undertanken to determine operating parameters of circular ring microstrip antenna elements. The approach taken was to express the electromagnetic fields within the antenna in terms of cylindrical modes and to employ an admittance boundary

condition at the radiating aperture. A ring microstrip is a radiating element which may be used in conjunction with other elements, either microstrip or waveguide, to provide a compact dual-band antenna system. The results of this study showed that ring microstrip antennas are very practical and useful devices. Sufficient material may be removed from the center of the microstrip disk antenna to allow the insertion of a high frequency element. Also substantial reduction of the operating frequency for a microstrip antenna of given maximum radius can be achieved. This property is useful for arrays of microstrip antennas since the seperation between elements may be increased because of the smaller microstrip element. More detailed results of this investiation are given in appendix 2.

LIST OF MANUSCRIPTS SUBMITTED OR PUBLISHED UNDER ARO SPONSORSHIP:

1. C. M. Krowne and A. R. Sindoris,"H-Plane Coupling Between Rectangular Microstrip Antennas," Printed Circuit Antenna Technology Workshop, New Mexico State University, Las Cruces, NM, October 1979.

2. C. M. Krowne and A. R. Sindoris,"H-Plane Coupling Between Rectangular Microstrip Antennas," Electronics Letters, vol. 16, no. 16, pp. 211-213, 13 March 1980.

3. C. M. Krowne and A. R. Sindoris,"Calculation of H-Plane Mutual Coupling Between Rectangular Microstrip Antennas," IEEE International Symposium on Antennas and Propagation, Quebec, Canada, June 1980.

4. J. W. Mink,"Circular Ring Microstrip Antenna Elements," IEEE International Symposum on Antennas and Propagation, Quebec, Canada, June 1980.

5. C. M. Krowne and A. R. Sindoris,"Input-Impedance Smith Chart Curves for H-Plane Mutual Coupling Between Two Rectangular Microstrip Antennas," Submitted to IEEE International Sysposium on Antennas and Propagation, Los Angeles, CA, June 1981.

SCIENTIFIC PERSONNEL SUPPORTED BY THIS PROJECT AND DEGREES AWARDED:

Mr. David Shain	M.S. EE
Mr. Floyd Blackwell	B.S. EE
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APPENDIX 1

INPUT-IMPEDANCE SMITH CHART CURVES FOR H-PLANE MUTUAL COUPLING BETWEEN TWO RECTANGULAR MICROSTRIP ANTENNAS

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Introduction

Non-radiative coupling between microstrip antenna elements can be especially important in arrays where the separation between elements is on the order of a few wavelengths λ.[1] At 1.5 GHz, λ=6.7-12 cm in materials with relative dielectric constants of polystyrene (ϵ_r=2.56) through ϵ_r=10. Here we extend earlier work[2] where this coupling is calculated for two closely spaced rectangular microstrip antennas in H-plane alignment. Extensive results are presented for the antennas in H-plane orientation with the dielectric substrate below the antennas ranging from ϵ_r=2.5 to ϵ_r=10.

The two coupled rectangular antennas are shown in Figure 1. Microstrip feed lines are seen to the left of each of the antenna patches. The top antenna is fed with a quasi-TEM wave mode which is assumed to propagate in the same direction as the microstrip feed line. This direction is parallel to both antenna edges where mutual coupling is largest. The other microstrip feed line returns the inputted energy, again propagating in the quasi-TEM mode. This output line is returned to ground through a load Z_L.

Theory

Considering only the antenna patches, they can be modeled using parallel-coupled transmission line theory. The impedance matrix relating the voltage and current at the four antenna ends is[3]

$$\begin{bmatrix} V_o \\ V_i \\ V_t \\ V_b \end{bmatrix} = \begin{bmatrix} z_{oo} & z_{oi} & z_{ot} & z_{ob} \\ z_{oi} & z_{oo} & z_{ob} & z_{ot} \\ z_{ot} & z_{ob} & z_{oo} & z_{oi} \\ z_{ob} & z_{ot} & z_{oi} & z_{oo} \end{bmatrix} \begin{bmatrix} I_o \\ I_i \\ I_t \\ I_b \end{bmatrix} \tag{1}$$

where the subscripts o, i, t, and b denote output, input, antenna end opposite input (top of Figure 1), and antenna end opposite output (bottom of Figure 1). The antenna ends are made radiating by adding to each a parallel radiation admittance Y to ground. On the right-hand antenna ends, adding these admittances imposes the conditions

$$I_t = -Y\,V_t \tag{2a}$$

$$I_b = -Y\,V_b\ . \tag{2b}$$

Combining Eqs. (1) and (2) yields the 2-port

$$\begin{bmatrix} V_o \\ V_i \end{bmatrix} = [Z'] \begin{bmatrix} I_o \\ I_i \end{bmatrix} \tag{3}$$

where

$$z_{kj}' = z_{kj} - K^{-1}Y[B_j z_{ot} + A_j z_{ob}] \tag{4a}$$

$$B_o = z_{ot} + Y[z_{oo}z_{ot} - z_{oi}z_{ob}] \tag{4b}$$

$$B_i = z_{ob} + Y[z_{oo}z_{ob} - z_{oi}z_{ot}] \tag{4c}$$

$$K = (1 + Yz_{oo})^2 - (Yz_{oi})^2 \tag{4d}$$

$$A_o = B_i\ ,\ A_i = B_o\ ;\ k,j = o \text{ or } i. \tag{4e}$$

The input impedance $Z_{in} = V_i/(I_i + V_iY)$ can be found by using the condition

$$I_o = -(Y+Y_L)V_o \tag{5}$$

and the elements of the following ABCD matrix (found by rearranging Eq. 3):

$$\begin{bmatrix} V_i \\ I_i \end{bmatrix} = \begin{bmatrix} a & b \\ c & d \end{bmatrix} \begin{bmatrix} V_o \\ I_o \end{bmatrix} \tag{6}$$

Here

$$a = z_{ii}'/z_{oi}'\ ;\ b = z_{io}'-(z_{oo}'-z_{ii}')/z_{oi}'\ ;\ c=1/z_{oi}'\ ;\ d=z_{oo}'/z_{oi}'\ . \tag{7}$$

Thus

$$Z_{in} = \left[\frac{-c + d(Y+Y_L)}{-a + b(Y+Y_L)} + Y\right]^{-1}\ . \tag{8}$$

The extreme simplicity of Eqs. (1)-(8) is due to both the reciprocity of passive networks and the geometric symmetry of the coupled antennas.

Results

Figures 2-4 plot Z_{in} on Smith Charts for the three cases $\epsilon_r = 2.5$, 5, and 10 (Z_L=50Ω). Consider Fig. 2. Four antenna patch widths w are plotted: w = 15,5,2.5, and 0.5 cm going from left to right. For each width, two separation values between the antenna patches are used to compute Z_{in}; they are s = 0.1587 and 5.0 cm going from left to right. For all figures, the antenna patch lengths ℓ are held at 5.0 cm. The frequency f increases in a clockwise manner. f points are labeled only at the beginning and end of each curve, with the step size given in parentheses next to the beginning f value. Notice that the effect of s on Z_{in} in going from ϵ_r=2.5 to 10 (Figs. 2 to 4) progressively increases, in accordance with the increasing non-radiative near-field coupling.

Acknowledgment

One of the authors (CMK) would like to thank Kuang-Yuh Wu for computer programming assistance.

Summary

The effect of mutual H-plane coupling on the input impedance of a rectangular microstrip antenna has been determined in a comprehensive set of curves. These curves are based on a particular choice[3] for the radiation admittance, assumed identical for all four antenna patch ends. A different choice for Y is not expected to change the fundamental behavior displayed by the impedance curves.

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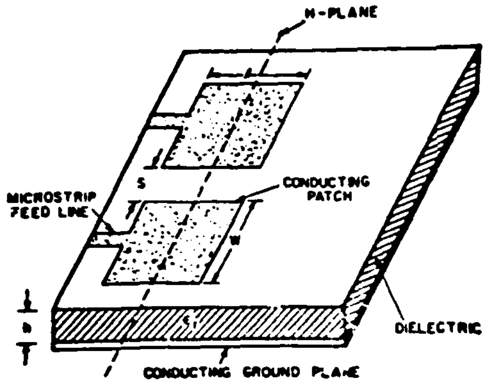

Fig. 1 Perspective drawing of two H-plane coupled rectangular microstrip antennas.

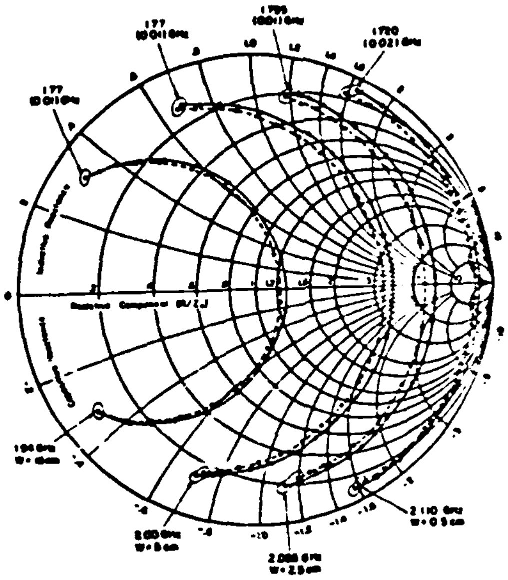

Fig. 2 Z_{in} versus w and s(s=0.1587 cm, ----; s=5 cm, ——). ϵ_r=2.5.

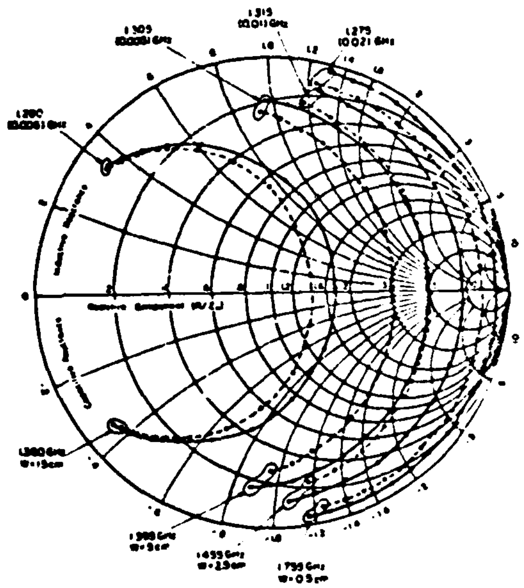

Fig. 3 Z_{in} versus w and s. ϵ_r=5.0.

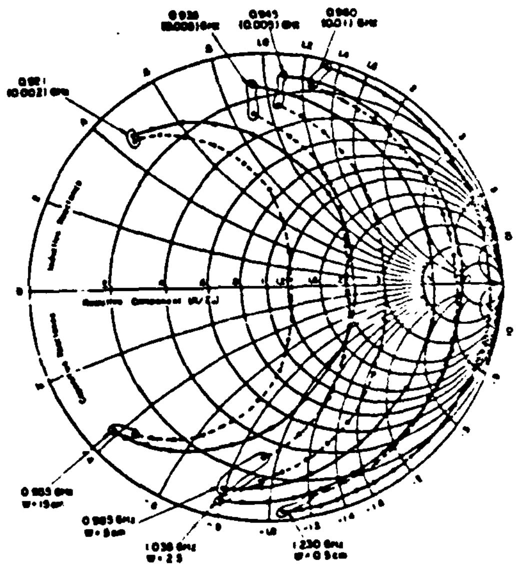

Fig. 4 Z_{in} versus w and s. ϵ_r=10.0.

APPENDIX 2

CIRCULAR RING MICROSTRIP ANTENNA ELEMENTS

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INTRODUCTION

As microstrip antennas [1] find more sophisticated applications, simple rectangular or circular geometries are no longer adequate. A ring microstrip is an example of a radiating element which when employed in conjunction with other elements may provide a compact dual-band antenna system. The ring geometry, shown in Fig 1, may be combined with a higher frequency radiating element within its aperture. A second microstrip antenna or a conventional waveguide element which radiates through the aperture after the ground plane and dielectric layer is removed from the aperture region may be employed as the high frequency element. The rectangular ring microstrip antenna was first proposed by and experimentally investigated by Kerr [2] as the low frequency element of a dual-band feed for a parabolic reflector antenna.

Since, for a given maximum size the operating frequency may be substantially lowered, these elements will also find application in phased arrays. This property allows one to construct a more dense array of elements thereby reducing the grating lobe problem. The purpose of this paper is to provide an analytical understanding of microstrip ring antennas along with experimental confirmation.

THEORY

Since cylindrical geometry is employed and the separation between the antenna and the ground plane is small in terms of wavelength, the fields within the region between the circular conducting ring and the ground plane may be expressed in the following form:

$$E_z = E_0\left[J_n(k\rho) + A\,N_n(k\rho)\right]\cos(n\varphi)$$

$$H_\rho = -\frac{j\omega\epsilon n}{k^2\rho}E_0\left[J_n(k\rho) + A\,N_n(k\rho)\right]\sin(n\varphi)$$

$$H_\varphi = -\frac{j\omega\epsilon n}{k}E_0\left[J_n'(k\rho) + A\,N_n'(k\rho)\right]\cos(n\varphi)$$

Where $J_n(\chi)$ and $N_n(\chi)$ are Bessel functions of the first and second kind respectively, the constant A is chosen to satisfy the boundary conditions. A relationship which

couples the interior fields to the exterior fields must now be developed. The approach taken here is to lump all external stored energy and radiation losses through a complex wall admittance, Y_a and Y_R, at r equal a and R respectively [3]. The boundary conditions are then obtained through application of transverse resonant techniques [4]. Thus, one may write the boundary conditions as:

$$\begin{bmatrix} \left[J_n'(kR) + jY_R Z_c J_n(kR)\right] & \left[N_n'(kR) + jY_R Z_c N_n(kR)\right] \\ \left[J_n'(ka) - jY_a Z_c J_n(ka)\right] & \left[N_n'(ka) - jY_a Z_c N_n(ka)\right] \end{bmatrix} \begin{bmatrix} 1 \\ A \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \end{bmatrix}$$

A non-trivial solution to this system of equations is obtained when the determinate of the coefficients equals zero. Because the wall admittances are complex the wave number k must also be complex; however, since microstrip antennas are relatively high Q devices the real part of k will be large compared to the imagionary part. To facilitate the computation of the Bessel functions with complex arguments, they were expanded in a Taylor's series about the real part of their argument. To obtain the complex resonant frequency, the transcendental equations which result from application of the boundary conditions were solved numerically using the Gauss-Seidel [5] method of successive displacements.

COMPUTED RESULTS

The frequency characteristics of circular ring microstrip antennas were obtained as a function of relative aperture radius. In all cases the resonant frequency was normalized to that of an ideal disk microstrip antenna with the same radius, and assumed magnetic wall boundary conditions. Since, at this time, an exact value for the complex wall admittance is not known, an equivalent admittance was derived for each of several commonly used models [6, 7, 8]. This admittance was then used as the boundary condition for the computation of the resonant frequency as a function of relative aperture radius. Results of these calculations are shown in Figure 2; also shown in Figure 2 is the comparison with experiment. Figure 3 shows the variation in resonant frequency as the substrate thickness and relative aperture is varied using as the wall admittance the value obtained for a slot in a screen [6]. Typical field distributions for the fields

within the region between the conducting ring and the ground plane will be presented.

CONCLUSION

The results of this study show that ring microstrip antennas are very practical and useful devices. Sufficient material may be removed from the center of a microstrip disk antenna to allow the insertion of a high frequency element in that region, thus, making dual frequency operation feasible. Also a substantial reduction of the operating frequency for a microstrip antenna of given maximum radius can be achieved. This property is useful for arrays of microstrip antennas. Preliminary results also show an increase of bandwidth as the relative aperture radius is increased.

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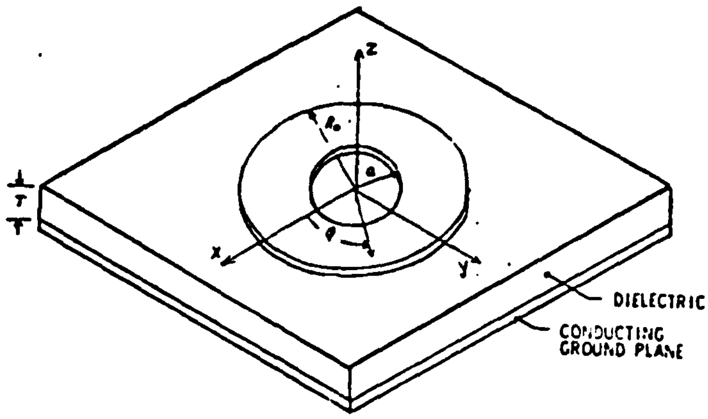

FIGURE 1 CIRCULAR RING STRUCTURE

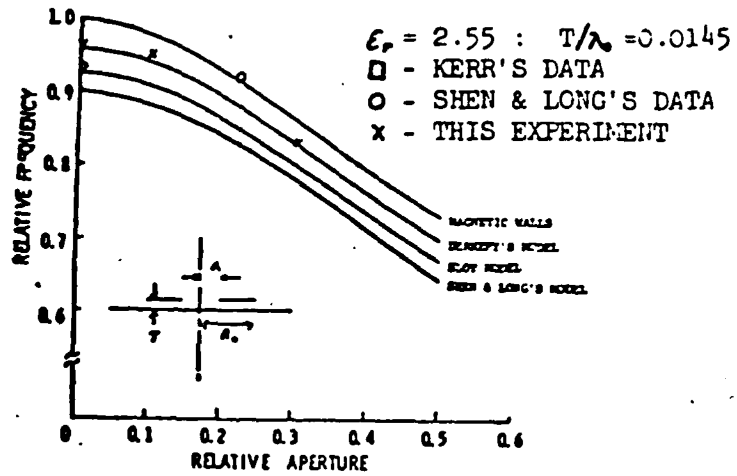

FIGURE 2 CHANGE OF RESONANT FREQUENCY DUE TO AN APERTURE AND EDGE ADMITTANCE

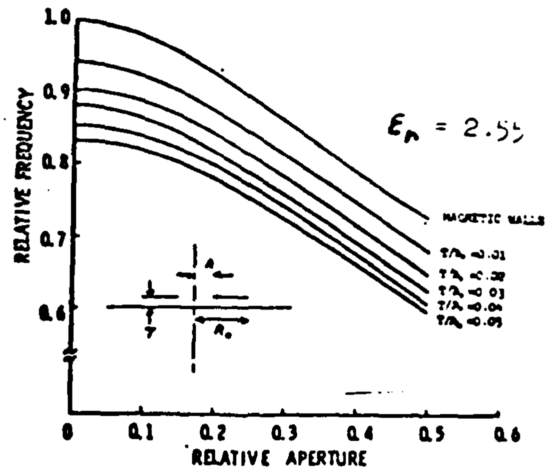

FIGURE 3 CHANGE OF RESONANT FREQUENCY DUE TO AN APERTURE AND DIELECTRIC THICKNESS